Phased array system design

Shannon Wanner

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Phased array system design

by

Shannon Wanner

A dissertation submitted to the graduate faculty
in partial fulfillment of the requirements for the degree of

DOCTOR OF PHILOSOPHY

Major: Electrical Engineering

Program of Study Committee:
Robert J. Weber, Major Professor
Jiming Song
Degang Chen
Mani Mina
Stephen Vardermann

Iowa State University
Ames, Iowa
2008

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DEDICATIONS

I would like to dedicate this work to my parents for their enduring support and my mother for putting up with my intolerable questions when I was younger. I would like to thank my sister for encouraging me to take the path that is less traveled, the lord and my savior for guiding and protecting me on this path and also dedicate my work to my professor, Robert Weber.
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## GLOSSARY OF TERMS

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<th>Abbreviation</th>
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<tr>
<td>CPU</td>
<td>Computational programming unit</td>
</tr>
<tr>
<td>DSP</td>
<td>Digital signal processing</td>
</tr>
<tr>
<td>EBG</td>
<td>Electromagnetic band gap</td>
</tr>
<tr>
<td>FPGA</td>
<td>Field programmable gate array</td>
</tr>
<tr>
<td>FDTD</td>
<td>Finite difference and time domain</td>
</tr>
<tr>
<td>GL</td>
<td>Grating lobe</td>
</tr>
<tr>
<td>HFSS</td>
<td>A commercial finite element computational program</td>
</tr>
<tr>
<td>MIMO</td>
<td>Multiple inputs and multiple output communication system</td>
</tr>
<tr>
<td>PA</td>
<td>Power amplifier</td>
</tr>
<tr>
<td>PCB</td>
<td>Printed circuit board</td>
</tr>
<tr>
<td>PD</td>
<td>Phase detector</td>
</tr>
<tr>
<td>PLL</td>
<td>Phased locked loop</td>
</tr>
<tr>
<td>LAN</td>
<td>Local area network</td>
</tr>
<tr>
<td>LOS</td>
<td>Line-of-sight-path</td>
</tr>
<tr>
<td>NLOS</td>
<td>Non-line-of-sight-path</td>
</tr>
<tr>
<td>RF</td>
<td>Radio frequency</td>
</tr>
<tr>
<td>$S_{ij}$</td>
<td>Scattering parameter from port $i$ to port $j$</td>
</tr>
<tr>
<td>SNR</td>
<td>Signal-to-noise ratio</td>
</tr>
<tr>
<td>QAM</td>
<td>Quadrature amplitude modulation</td>
</tr>
<tr>
<td>QPSK</td>
<td>Quadrature phase shift keying</td>
</tr>
<tr>
<td>VCO</td>
<td>Voltage controlled oscillator</td>
</tr>
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ABSTRACT

This dissertation presents a new architecture for controlling the amplitude and phases of signals in a fully integrated phased array system. The phase shifting is performed at baseband and is up-converted to a RF band using a phased locked loop as a frequency synthesizer which allows for precise control of the RF phases. A structure for a phase aid is presented for a fast phase hopping system. A mathematical model was shown for an adjustable amplification within a phased locked loop to mitigate phase distortion. A mutual coupling correctional network was proposed to mitigate pattern distortion induced from the active impedance model. An automatic tuner to reduce manufacturing variations within an array was proposed that will reduce unwanted effects from the active impedance model and minimize the losses due to reflections. A polarization alignment system was also proposed that will help improve the overall gain between the transmitting and receiving system.
CHAPTER 1. BACKGROUND

1.1 Introduction

The phased array antenna system has many applications in wireless applications, especially MIMO (multiple inputs and multiple outputs) communication. By using multiple antennas to transmit and receive the signal, the transmit rate is pushed closer towards the channel capacity limit while simultaneously improving security [2]. Another application of such a system is in Sensor Array Networks where information from a single sensor can be directed or transmitted to a specific receiver by steering the antenna in the right direction [3]. Since the transmitted signal can be steered to a specific receiver and nulled in other directions, the security of the signal can be improved. A phased array antenna system can be utilized by the military to transmit and receive secure information [2]. It also has applications in mobile LANs, adaptive dynamic array processing for antennas and automotive radars for collision control, path/lane control, etc [4], [5]. Phased array applications have been around since the early 1940’s when first developed with military intentions. The primary purpose of a phased array is to agilely steer the main beam of an antenna in a given angular direction of the line-of-sight-path (LOS) which is labeled as the direct path in Fig. 1.1. This will enhance the reception of the direct path signal and reduce the reception of non-line-of-sight (NLOS) components which is labeled as the reflected paths in Figure 1.1: Illustration of Multi-path fading in a communication link [1].
Figure 1.1. As a result of steering, the multipath effects of the communication link will be reduced. The reduction of multipath fading will inherently increase the channel capacity of the communication link. A demonstration of steering by a phased array is illustrated in Fig 1.2. This is accomplished by setting phases and amplitudes on each element of the array. A secondary aspect of phased array antennas is the ability to add a null in the direction of an unwanted signal, for instance an enemy trying to jam the communication system or a

Figure 1.2: The steering of the main beam of antenna at 0, 90, and 180 degrees for part A, B, and C respectively. The simulation was done using Artdtool for Matlab with a 10 element, omni-directional, ordinary end fire array spaced a half wavelength apart [6].
A hacker trying to gain access to a private network [2]. A majority of the work done in the early development of phased array has been performed in the military settings thus limiting the commercial development of the area. The primary cause of this limitation to military rather than the commercial environment was due to the cost and accurate characterization of the components. Today’s technology has matured significantly enough to allow for the technology to break into the commercial world. The accuracy of simulations has been greatly improved through both commercial and proprietary computational electromagnetic programs. The limitations of these programs are primarily due to the requirement of a high-end CPU to

![VARIOUS SPATIAL FILTERING WEIGHTINGS](image)

Figure 1.3: Part A. shows the response of a filter that corresponds to an equal amplitude phased array, part B shows the response of a filter that corresponds to a binomial weighted phased array and part C shows the response of a Tchebyscheff weighted phased array. The simulation was done using Arqtool for Matlab with a 10 element, omni-directional, ordinary end fire array [6].
produce an accurate simulation which increases with the size of the grid. Integration of the RF components onto a silicon chip has reduced the variability of the amplitude and phase characteristics of the devices and has significantly driven the cost down. Beam formation algorithms have in the past been developed using spatial filter theory and these have been thoroughly studied. Several responses of different spatial filters are shown in Fig. 1.3. Similar tradeoffs between stop band and pass band characteristics in a circuit filter design can be observed in an array such as shown in Fig 1.3. There will be a tradeoff between directivity.

Figure 1.4: Results of optimization routines for beam formation in the direction of line sight for incoming signals at $\theta = \pi / 6$ and $\theta = 2\pi / 3$ with 4 omni-directional elements spaced 0.075 m apart on the x-axis at 2 GHz. Part A shows the results of the ESPRIT algorithm, part B shows the result of the Temporal Reference algorithm, and Part C shows the result of the MUSIC algorithm using the Smart Antenna System for Mobile Communication [7].
in the pass band and side lobe ripples in the stop band. An inherent drawback of this theory is accurate adjustability of the phase and an amplitude characteristic is required for each elemental excitation of the phased array. It is also extremely complex to produce an optimal design in a large phased array in a non-static environment. Current research in beam formation is producing designs that adaptively converge to an optimal solution using optimization techniques such as convex, genetic, linear, and gradient optimization algorithms. Responses from MUSIC, Temporal Reference, ESPIRIT algorithms are shown in Fig. 1.4. There will be a tradeoff between the ability of the algorithm to identify the direction of the incoming signal within a pass band and rejection of other signals within the stop band. This can be observed in part A and part B of Fig 1.4. The algorithm in Part C. of Fig 1.4 is only used to identify the direction of the incoming signals. This information can then be utilized to design an appropriate filter.

1.1.1 Hypothesis

An integrated phased array system can be built using shift registers and phase locked loops, to control the phase of the electronics with arbitrary accuracy. This will in turn allow the direction of the beam pattern of the main beam of the array to be digitally controlled and allow nulls to be set in a predictable fashion.

1.2 Approach

A three by three phased array was built and tested. An FPGA was used to perform the digital circuitry tasks and a phased locked loop was used as an RF up-converter to control the main beam of the antenna in a predictable manner. The system was measured in an anechoic
chamber in order to avoid interference, reflections, and to allow for precise measurement of the beam pattern in the far field measurement location.

1.3 Summary

The hardware platform, which was outlined in Sec. 1.2, would enhance the current state of the art by allowing more accurate phased array systems to be built with less expense. This in turn allows the field of space-time adaptive communications and MIMO systems to break the barrier of moving beyond theoretical ideas into product development. The need for such systems will become more critical as demands for wireless bandwidth increases worldwide. Better control of beam patterns is already becoming critical in densely populated areas such as European countries.
CHAPTER 2. REVIEW OF LITERATURE

2.1 Introduction to Phase Shifters

A phase adjusting circuit (phase shifter, etc.) is absolutely critical in a phased array system. The evaluation of phase shifters is most often evaluated in terms of cost, linearity, and available phase shift. Phase shifters can be broken down into three different categories: time delay element, filter element, or active devices.

2.1.1 Time Delay

Time delay elements have often been reported using microstrip lines [8]. An inherent drawback of a time delay element is that most often they are a non-variable phase shifter unless switches are implemented and then the number of different phase shifts available is directly related to the number of switches, thus increasing the overall cost of the system. Another drawback of using transmission lines as a phase shifter is that the size is inversely proportional to frequency and most commercial applications are implemented below 10 GHz where transmission lines are physically large. The shift in time, $T_{\text{delay}}$, of a transmission line is related to the phase and the frequency of operation, $\omega_{\text{rf}}$, as shown below

$$\theta_{\text{shift}} = \omega_{\text{rf}} T_{\text{delay}},$$

(2.1)

The resultant output of the shift can then be represented as

$$\theta_{\text{shift}} = \omega_{\text{rf}} T_{\text{delay}},$$

(2.2)
\[ V_{\text{out}} = V_{\text{in}} \sin(\omega_{rf} t + \theta_{\text{shift}}). \]  \hspace{1cm} (2.3)

Phase shifters implemented using filters are most often done using a variable reactance element such as a varactor [9], or ferrite based device [10], thus making the transfer function variable. These have been implemented at baseband, RF, and directly at the antenna’s terminals. In most cases, the inherent drawback of variable phase shifters lies in the drift of the phase shift and the linearity of the phase shifter. The phase shift of a filter can be determined by its phase transfer function,

\[ \theta_{\text{shift}} = \tan^{-1}\left( \frac{\text{imag}(H(\omega))}{\text{real}(H(\omega))} \right), \]  \hspace{1cm} (2.4)

where \( H(\omega) \) is the transfer function of the filter. The magnitude of the transfer function will also vary when adjusting the reactive elements of the filter. This will cause a coupling between the amplitude and phase of the excitation and produce a distortion in the desired array pattern as the filter is varied to steer the array.

### 2.1.2 Active Phase Shift

Active phase shifters have been produced by using injection locking properties of voltage controlled oscillators [11] or phased locked loops [12]. In the past, the inherent drawback of phased locked loops has been the cost of their implementation. This has been remedied, however, by the decreasing cost of integrated phased locked loops and the increasing availability of phased locked loops that operate above 1 GHz. This availability of phased locked loops above 1 GHz will likely continue until Moore’s Law ceases to be true. An inherent gain of the phased locked loop implementation is the ability to perform a base
band phase shift. That is typically performed at a lower cost solution than with a RF shifter. The inherent drawback of injection locking is that the signals on the antenna elements are coupled to their nearest neighbors and this reduces the degrees of freedom allowed for beam formation although it is sufficient enough for crude scanning.

2.2 Introduction to Antenna, Arrays, and Mutual Coupling

Antennas have progressed from simple radiating elements, such as a simple monopole or a loop antenna, to a more integrated approach such as using a microstrip antenna or an inverted F antenna. This integration allows the feasibility of building large arrays with super-directivity properties; however, a large inherent drawback of large arrays is increased mutual coupling that will distort the field pattern [13]. The increase in mutual coupling has led to an increase in methods to deal with or reduce the mutual coupling between elements.

2.2.1 Antenna

In modern day consumer electronics, there are four different types of antennas in use. These are the inverted F, the microstrip antenna, the helix antenna and the dipole antenna. The selection criteria for any antenna can be reduced down to its polarization, manufacturability, bandwidth, and field pattern. The manufacturability considerations include cost of material, robustness against manufacturing errors, mutual coupling in the array, and integration of the components. Field pattern considerations can be broken down into two categories - bandwidth and directivity of the antenna. The application dictates which property is more desirable. Dipole antenna’s advantages are H-plane omni-directional field patterns, a medium wide bandwidth, and medium manufacturing complexity [14]. A
microstrip antenna’s advantages are ease of manufacturing, dual polarization capabilities, and ultra-wide bandwidth capabilities [14]. An inverted F antenna’s advantages are an approximate omni-directional pattern, ease of manufacturing, and medium width bandwidth capabilities. A helical antenna’s advantages are an H-plane omni-directional field pattern capability, a dual polarization capability, and a high bandwidth [14]. Typical criteria for an antenna in a phased array system are directivity, dual polarization support, and ease of manufacturability. Therefore, the microstrip antenna often seems to be the most suitable antenna. There are numerous variants of microstrip antennas. The selection criterion depends on polarization capability and gain.

### 2.2.2 Array Design

Previous work [14], not including mutual coupling effects, suggest that any antenna array can be factored into two components. One component, $E_{element}$, contains the field pattern of a single radiating element and the second component, $AF$, that contains geometric information concerning how the antennas are arranged [14]

$$E_{total-field} = \sum_{n=1}^{N} E_{element}^n = E_{element} AF. \quad (2.5)$$

The importance of this separation of terms is the fact that the arrays pass band and stop band can be designed by controlling both the amplitudes and phases of the array factor. The element field pattern then only affects the angle at which the pass band can be placed and the directivity of the overall array pattern. Consider an array with M elements in the x-axis direction and N elements in the y-axis direction, assuming then that the antennas are
positioned linearly about the x-axis and y-axis, and M and N are odd, the array factor can then be defined as

\[ AF = \sum_{m=-(M-1)/2}^{(M-1)/2} e^{jkd_m \sin(\theta) \cos(\phi)} \sum_{n=-(N-1)/2}^{(N-1)/2} |A_{nm}| e^{-j\theta_{nm} e^{jkd_n \sin(\theta) \sin(\phi)}} \]  

(2.6)

where \(d\) is the center-to-center spacing of the elements. The array factor is solely determined by the amplitude, \(|A_{nm}|\), and phase, \(\theta_{nm}\), of the excitation placed on each element. The array factors for other geometries, such as circular, rectangular, etc., arrays have also been formulated [16] [17]. An excitation can be defined as the following

\[ A_{nm} = |A_{nm}| e^{-jk_{\text{mm}} \psi_o}. \]  

(2.7)
The position vector, \( r_{nm} \), is dependent upon the location of the patch, which is defined in Fig 2.1. The main beam of the antenna can be steered in the direction of \( (\theta_o, \phi) \) by setting the phases on the antenna element where the resulting variable \( \hat{\psi}_o \) is given by [17]

\[
\hat{\psi}_o = \sin \theta_o \cos \phi_o \hat{a}_x + \sin \theta_o \sin \phi_o \hat{a}_y + \cos \theta_o \hat{a}_z. \tag{2.8}
\]

The vector, \( \hat{\psi}_o \), is defined from the center of the array to center of the main lobe. This results in a constant beam direction due to the fact that the corresponding exponential terms in the array factor are canceled [15]. One concern in the design of a phased array is the appearance of grating lobes. A grating lobe is the duplication of the main lobe at a different angle. An example of a grating lobe can be observed in Part B of Fig 1.2. Gating lobes can appear under the following conditions

\[
2 \pi \frac{d}{\lambda} \left( \sin \theta - \sin \theta_o \right) = 2\pi n,
\]

where \( n \) is an integer. The grating lobe appears at an angle, \( \theta_{GL} \), determined by [15]

\[
\sin \theta_{GL} = \sin \theta_o + \frac{n\lambda}{d}. \tag{2.10}
\]

with

\[
\left| \sin \theta_{GL} \right| \leq 1.
\]

Without an appearance of a grating lobe, for a given scan angle \( \theta_o \) at a frequency \( f \), the maximum element spacing is given by [15]

\[
\frac{d}{\lambda} = \frac{1}{1 + \sin \theta_o}. \tag{2.12}
\]

Typical element spacing for patch array antennas is given by [15]

\[
0.5\lambda \leq d \leq \lambda. \tag{2.13}
\]
The closer the antennas are to each other, the more the mutual coupling increases. The further the antennas are apart the more that grating lobes are introduced. The appearance of grating lobes is similar to aliasing of a sampled signal as derived in signal processing theory. The placement of an antenna is effectively sampling in the spatial domain. In signal processing, one must sample the signal at a rate of twice the highest frequency component in order to avoid aliasing of the signal. Similarly, one must sample the spatial domain at a rate that is twice the wavelength of operation in order to avoid the appearance of grating lobes. The relationship between distance and mutual coupling is shown in Fig. 2.2. Mutual coupling in dB’s can be seen as a function of distance, relative dielectric, height of PCB, and frequency of operation [18].
2.2.3 Introduction to Mutual Coupling

The most commonly used method for calculating mutual coupling is through the spectral domain approach. This methodology, first proposed by Pozar, accounts for spatial, reflective, and surface wave coupling [19-20]. These coupling waves are as shown in Fig. 2.3. The type of coupling that is prevalent determines the hardware mechanism that is available to reduce the coupling. Pozar introduced the idea of the active impedance which states that the input impedance of one antenna element is affected by the coupling of the output of another antenna element [13]. The effects in a phased array design were explored by Dodov, who found that surface wave and reflective coupling are only prominent in long arrays and also stated the conditions for excitation of such waves [18]. The Floquet model was introduced to predict scan blindness in an infinite array [21].

Fig 2.3: A diagram showing the various methods that a field emanated from one antenna element interacts with neighboring antenna elements [18].
In a coherent system, the reflection coefficient is dependent upon the neighboring antennas. Assuming coherency, mutual coupling effects, to a first order approximation, can be described in terms of an active reflection coefficient as can be derived from the flow diagram in Fig. 2.4 and given below [13]

\[
\Gamma_1 = \frac{V_{\text{refl}}}{V_{\text{trans}}} = \frac{A_1 S_{11} + A_2 S_{12}}{A_1} = S_{11} + \frac{A_2}{A_1} S_{12},
\]

(2.14)

\[
\Gamma_2 = \frac{V_{\text{refl}}}{V_{\text{trans}}} = \frac{A_2 S_{22} + A_1 S_{21}}{A_2} = S_{22} + \frac{A_1}{A_2} S_{21}.
\]

(2.15)

The field pattern of a two element array shown in Fig. 2.5 can be described in terms of mutual coupling parameters as

\[
E_\phi \propto E_{\text{element}} \left\{ (A_1 + A_1 S_{11} + A_2 S_{12}) e^{i (kd/2) \sin \phi} ight. \\
\left. + (A_2 + A_2 S_{22} + A_1 S_{21}) e^{i (-kd/2) \sin \phi} \right\},
\]

(2.16)
where $A_1$ and $A_2$ are the complex feed excitations and $E_{\text{element}}$ is the field pattern emitted by a single antenna.

### 2.2.4 Mutual Coupling Reduction

There have been two different methods used to reduce mutual coupling effects in arrays, electromagnetic band gap structures [23] and perforated ground planes [24]. Electromagnetic band gap structures (EBG) and a perforated ground plane shown in Fig. 2.6, allow for the physical reduction of coupling due to surface waves by removing the electromagnetic boundary conditions which allow them to propagate. These techniques have been shown to reduce mutual coupling up to -15 dB. The disadvantage of an EBG and a perforated ground plane is that they do not reduce coupling through the air or through reflection and therefore must be combined with the predistortion technique in order to get an undistorted field. A disadvantage of pre-distortion techniques is in the fact that the technique must be adaptive due to environmental changes that will affect coupling parameters. The EBG and the perforated ground plane are passive structures.

Fig 2.6: A perforated ground plane (Left) and an electromagnetic band gap (Right) [24], [23].
2.3 Phased locked loops, Auto-tuners, and Adjustable Amplifiers

Phased locked loops have progressed from a purely analog form to a mostly digital approach. The integration of components in a phased locked loop has almost reached full potential. The filter is still left external from the VLSI chip to allow for adaptability of the phased locked loop to a wide variety of applications. Auto-tuners have been recently proposed in the last couple of decades [28]. The reason that they are not being implemented is primarily due to the non-linear nature of a varactor or pin diode. These devices will generate harmonic content that couples into the antenna when large output signals are induced from a power amplifier. Adjustable gain amplifiers have been developed using pin diodes as attenuators due to the low cost of components and ease of implementation. However, amplifiers using adjustable gain do exist and utilize current versus gain more efficiently. The gain of the amplifier can be written as

\[ Gain = g_m R_{out} \]  

(2.17)

<table>
<thead>
<tr>
<th></th>
<th>Type 1 Filter</th>
<th>Type 2 Filter</th>
<th>Type 3 Filter</th>
</tr>
</thead>
<tbody>
<tr>
<td>Phase position</td>
<td>Zero</td>
<td>Zero</td>
<td>Zero</td>
</tr>
<tr>
<td>Phase velocity</td>
<td>Constant</td>
<td>Zero</td>
<td>Zero</td>
</tr>
<tr>
<td>Phase acceleration</td>
<td>Increasing</td>
<td>Constant</td>
<td>Zero</td>
</tr>
</tbody>
</table>

Table 2.1: Phase error resulting from different types of phase inputs [26].
in terms of its input transconductance, $g_m$, and the output impedance, $R_{out}$, where $g_m$ is dependent on the device current. Therefore, the gain can be made variable by either adjusting the load impedance or bias current.

### 2.3.1 Phased Locked Loop and Acquisition Aids

The acquisition aids published to date include but are not too limited to an adjustable filter in the phased locked loop. The acquisition time required to lock in a signal can be found to be proportional to the bandwidth of the input filter. However, larger bandwidths will cause additional phase noise and spurious content to appear in the output signal. Therefore, Lui and Li purposed adjusting the bandwidths of the filters based on the magnitude of the error signal [25], thus, incorporating the best properties of high bandwidth and low bandwidth in the phased locked loop. The type of filter used depends on the type of input signal and the phase error that can be tolerated in the output signal, which can be determined from Table 2.1. Typical application of type 1 or type 2 filters would be in a frequency synthesizer where phase acceleration is kept at a minimum and a type 3 filter could be used in a frequency recovery system where phase acceleration is a major concern. There has been considerable research in the architecture of all three types of filters that will guarantee a stable system [27].

### 2.3.2 Adjustable Amplifier

Typically, even though adjustable gain provides higher degrees of freedom in terms of beam formation, its use in a phased array is limited due to the fact that adjusting the gain
of the amplifier or using a variable attenuator, will change the phase characteristics on each radiator. This effect can be minimized and be de-embedded from the system by using a look up table and through the use of pre-distorting.

2.3.3 Automatic Tuner

Previous work in developing automatic tuning systems for antennas consists primarily of one structure as shown in Fig. 2.7 [28]. The reflected energy from the antenna is sensed by a power detector through a coupler. The reflected energy can then be used as an objective function in conjunction with an optimization algorithm to tune the antenna. Advantages of the structure consist of minimizing loss and beam distortion within the system. Disadvantages include the harmonics induced due to a large signal characteristic at the output of the PA in the transmitting mode. These effects are less detrimental in a receiving mode. However, De Mingo, Valdovinos, Crespo, Navarro, and Garcia suggested using notch filters to deal with harmonic generation as has been similarly done for power amplifiers [28]. These authors investigated to a limited extent the properties of adaptive, deterministic, and hybrid tuning algorithms.

![Fig 2.7: An automatic tuner for controlling the manufacturing mismatches in the antenna][28]
### 2.3.4 Polarization Adjustment

There have been numerous works describing microstrip patches that can radiate in different kinds of polarization modes. However, only one paper was found that describes a method of changing antenna polarization deterministically. The approach is shown in Fig. 2.8. The advantages of such a structure are that the polarization of transmitting or receiving signals can be tuned to the incoming wave and thus maximizing reception. The disadvantage is that the diodes go non-linear due to large signal characteristics of the signal at the output of a transmitter. This is less detrimental in the receiving mode where the alignment can be more easily performed. The tuning of the polarization is performed through the use of a varactor. The varactor capacitance can be change to dictate whether the length or the width of the antenna is resonant. This in turn determines the polarization of the antenna.

![Diagram of Antenna and Varactor](image)

**Fig 2.8:** Capture range of an automatic tuner for various mismatch resonances for different types of algorithms [29]
2.4 Summary

All previous results were judged on implementation, cost effectiveness, and manufacturability following current and future industrial trends. The author acknowledges that future trends may be subject to change based on innovation but remains firm on his criteria. The antenna array design and exploration has been explored. Future work needs to be implemented in randomization algorithms for the antenna design. Mutual coupling reduction methods have been explored but improvements are needed to determine which methods could be combined with baseband pre-distortion to further mitigate the effects. Phased locked loop aids do exist but need improvement if a fast phase hopping system is to be implemented with increasing data rates. Adjustable amplifiers also exist but their phase distortion will cause undesirable effects in the phased array. Further reduction of coupling will always be a benefit. The area of automatic tuners has been explored but has not been widely utilized. More experimental data, algorithms, and hardware design in the presence of a large signal inputs and outputs need to be obtained. Automatic tuners may become outdated if broadband antennas are capable of being designed with randomization procedures but may still be utilized to fine tune broadband antennas. Polarization adjustments have been explored but not utilized. The current approach for dealing with mismatch of polarization is by using multiple antennas with varying polarization. The author believes that an automatic polarization adjustment may be a more optimal in terms of both performance and cost. In summary, there is room for enhancement and continuing work in the field of phased array systems.
CHAPTER 3. HARDWARE THEORY
AND DESIGN METHODOLOGY

3.1 Introduction

The phase shifter section describes a shift register to produce fixed time delays of the input signal and forms the original basis for this project. The time delay values can then be chosen through the use of a multiplexer to pick the appropriate delay value. A latch register is used at the output of the multiplexer to produce a synchronous system. A phased locked loop is then utilized to up-convert the baseband signal to RF. The phased locked loop section describes an adjustable amplifier that is used provide more degrees of freedom to a phased array, a phase aid to provide a fast phase hopping system, and an offset PLL to provide a scaled free version of the hardware phase shifter. These were all added to the project to deal with specific problems and to provide a more optimal design. The antenna section describes a horn antenna design, which was used in experimentation, as well as a microstrip antenna and design procedures for an array. In addition, the antenna section also describes mutual coupling effects with a corrective network, scan blindness effects, and a polarization alignment scheme proposed to increase the SNR of the system.
3.2 Phase Shifters

3.2.1 Time Delay

A single element hardware block diagram for a prototype variable phase shifter is shown in Fig. 3.1 [30]. A reference clock is divided down by 16 to provide a data source and the output signal is represented as

\[ V_{ref}(t) = \frac{\pi}{4} \sum_{n=1,3,5}^{\infty} \frac{1}{n} \sin \left( \frac{2\pi nt}{T} \right), \]  

where \( T \) is the time period. The shift registers are shifted at one sixteenth of the clock rate of the system. The shift register then contains 16 different delayed versions, sampled on the rising edge of the clock, \( \omega_r \), as shown in below equation

\[ V_{shifted}(t,i) = \frac{\pi}{4} \sum_{i=1,3,5}^{\infty} \frac{1}{i} \sin \left( i\omega_r t + \frac{2\pi t}{16} \right), \]
for $i = 1, 2, ..., 16$ [30]. The PLL (phased locked loop) locks into phase with the shifted data and provides a 2.425 GHz source that is represented as

$$
\phi^i_0 = \frac{N}{M} \frac{2\pi i}{16} + \phi_{\text{loop delay}}, \quad (3.3)
$$

where

$$
-\pi \leq \phi^i_0 \leq \pi \quad (3.4)
$$

and $\phi_{\text{loop delay}}$ is the phase delay in the feedback network for $i = 1, 2, ..., 16$ and where $M$ and $N$ are the frequency divide ratios of the reference and the RF signal of the phased locked loop respectively as shown in Fig. 3.2.

![Fig 3.2: A diagram showing the basic functional elements within a phased locked loop.](image)
The division ratio produces a scaled version of the phase offset of the reference at the system output. Scaling can be minimized or eliminated by using a frequency offset system. In order to ensure stability and zero steady state phase error during phase hops, the phased locked loop contains a third order loop filter [27]. The settling time of the phased locked loop used can be seen in Fig. 3.3. The response determines the maximum bit rate that can be transmitted through the communication. The signal is sent to a power amplifier whose desired load impedance is matched to the inactive input impedance of the antenna terminals. The input signal can be represented as

$$V_{\text{antenna},p}(t, i) = C \sin(\omega t + \phi'_0),$$  \hspace{1cm} (3.5)$$

with $\omega$ being the RF output frequency and $\phi'_0$ is the phase shift introduced by the PLL and shift register.

Fig 3.3: A simulated phase error plot for the PLL unit step phase response.
3.3 Phased locked loop Components

3.3.1 Adjustable Amplifier

As shown in Fig. 3.4, in order to mitigate any phase error that could be caused by an adjustable gain amplifier, the amplifier is placed within a phased locked loop that provides negative feedback to reduce error caused by adjusting the amplitude of the signal. The Laplace transform of the loop gain is given by

\[ G = K_p K_f K_o K_n / s. \]  \hspace{1cm} (3.6)

The dynamic phase error can be written as follows

\[ \theta_d(s) = \theta_1(s) - \theta_2(s), \]  \hspace{1cm} (3.7)

Where \( \theta_1 \) and \( \theta_2 \) are shown in Fig 3.4. The error can then be further manipulated using standard control theory in terms of an input phase change and the distortion phase change

\[ \theta'_d(s) = \frac{1}{1 + GH(s)} \theta_1(s) - \frac{K_n}{1 + GH(s)} \theta_{distortion}(s). \]  \hspace{1cm} (3.8)
If the filtering is designed properly, the steady state error should reduce to zero. The time taken for convergence will then dictate the maximum bit rate that can be passed through such a system. The output phase can be found from the input phase by the following

\[ \theta_e(s) = \frac{G(s)}{1 + GH(s)} \theta_i(s) + \frac{1}{1 + GH(s)} \theta_{\text{distortion}}(s). \tag{3.9} \]

This derivation is similar to the standard PLL response derivation with the exception of an additional transient term. The transient’s time constant will depend on the open loop transfer function. This distortion correction technique is very similar to a Cartesian Feedback loop that is used for correction of amplitude to phase modulation conversion [32]. In this project, a pin diode attenuator was proposed for the adjustable amplifier due to its cost effectiveness and ease of manufacturing. An adjustable gain amplifier may be more desirable due to its higher efficiency.

### 3.3.2 Offset PLL

The multiplicative phase factor, caused by digital division in the feedback loop, can be eliminated by using mixers to translate the RF signal to baseband as shown in Fig. 3.5. In steady state, the mixers will produce a constant phase error which will be uniform over all antenna elements. Due to the uniformity of the phase error among the antenna elements, the error can be factored out of the array equation and thus mitigating a possibility of distortion.
The phase error can be written as

$$\theta_e(s) = \theta_1(s) - \theta_2(s)$$  \hspace{1cm} (3.10)

where $\theta_1$, and $\theta_2$ are shown in Fig 3.5. The transfer function of the error can then be written as

$$\theta_e(s) = \frac{1}{1 + GH(s)} \theta_1(s) - \frac{1}{1 + GH(s)} \theta_{offset1}(s) - \frac{1}{1 + GH(s)} \theta_{offset2}(s).$$  \hspace{1cm} (3.11)

The output can then be related to the error by

$$\theta_e(s) = \theta_1(s) - [\theta_0(s) + \theta_{offset1}(s) + \theta_{offset2}(s)].$$  \hspace{1cm} (3.12)

In the steady state, the error will approach zero assuming that the order of the denominator is of a higher power than the numerator and the filter is stable.

Thus,

$$\theta_e(s = 0) = \theta_1(s) - [\theta_0(s) + \theta_{offset1}(s) + \theta_{offset2}(s)] = 0.$$  \hspace{1cm} (3.13)
\[ \theta_1 = \theta_o + \theta_{\text{offset1}} + \theta_{\text{offset2}}. \]  
\[ (3.14) \]

In order to ensure that no false locks occur due to harmonics, a frequency planning technique of the following frequencies was used

\[ k(m \omega_{\text{offset1}} + n \omega_{\text{VCO}}) + l \omega_{\text{offset2}}, \]  
\[ (3.15) \]

where \( k, l, m, \) and \( n \) are integers and \( \omega_{\text{VCO}}, \omega_{\text{offset1}}, \) and \( \omega_{\text{offset2}} \) are the output frequency of the VCO, and LO frequencies for the offset mixers respectively. Adequate filtering was then applied to the corresponding harmonics.

![Diagram of a phased locked loop with an acquisition aid element inserted behind the VCO.]

Fig 3.6: A diagram of a phased locked loop with an acquisition aid element inserted behind the VCO.
3.3.3 Phase Aid

Phase aids can be introduced into a phased locked loop to reduce the transient time needed for convergence. The proposed structure shown in Fig. 3.6 is used to expedite convergence by summing a small delta phase error in the loop. The loop gain of the system shown in Fig 3.6 can be written as

\[ G = K_p K_j K_o K_n / s. \]  

(3.16)

The phase error can be written as

\[ \theta_e(s) = \theta_1(s) - \theta_2(s). \]  

(3.17)

where \( \theta_1 \) and \( \theta_2 \) are as shown in Fig 3.6. The equation can be further manipulated to produce the following

\[ \theta_e(s) = \frac{1}{1 + GH(s)} \theta_1(s) - \frac{K_o K_n}{s(1 + GH(s))} \theta_{offset}(s). \]  

(3.18)

The error can then be related to the output by the following equation

\[ \theta_o(s) = GH(s) \frac{\theta_e(s)}{K_n} + \frac{K_o}{s} \theta_{offset}(s). \]  

(3.19)

The additional offset term leads to an extra transient term

\[ \theta_o(s) = \frac{GH(s)}{1 + GH(s)} \frac{\theta_1(s)}{K_n} + \frac{K_o}{s(1 + GH(s))} \theta_{offset}(s). \]  

(3.20)
This extra transient can be formed such that when in combination with the original response it produces a pseudo convergence of the loop.

### 3.4 Antennas

#### 3.4.1 Antenna

The rectangular microstrip patch was chosen due to its popularity in an integrated transceiver, its capability of supporting dual polarizations, the ease of its manufacture, and having a well known approximate solution for its far field pattern. An example is shown in Fig. 3.7 [14]. The basic design of the microstrip antenna can be done in a five step process when the microstrip is treated as a resonant cavity. The assumptions for this model consist of enclosing the microstrip patch within a perfect magnetic conductor (PMC) artificially extending the length of the cavity to

![Diagram indicating length extension due to a fringing effect of the electric field and the corresponding dimension of a microstrip patch](image)

Fig 3.7: A diagram indicating length extension due to a fringing effect of the electric field and the corresponding dimension of a microstrip patch [34]
account for fringing effects, having an infinite ground plane, and assuming a dielectric, $\varepsilon_r$, exist only under the patch. The benefit of using the cavity model is that the main field lobe pattern is closely approximated. The disadvantages for the model include diffraction caused by the dielectric and finite ground plane that gives discrepancies for the field pattern for angles near grazing and behind the ground plane [33]. The dimensions can be calculated by the following sequence of steps.

1. Determine the [14]

$$W = \frac{1}{2} \frac{1}{f_r \sqrt{\mu_0 \varepsilon_0 \sqrt{\varepsilon_r + 1}}}.$$  \hspace{1cm} (3.21)

2. Calculate the length extension due to fringing effects caused by the electrical field extending past the microstrip patch as shown in Fig. 3.7. The effective dielectric constant due to the air and dielectric combined can be approximated by [14]

$$\varepsilon_{reff} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left[ \frac{h}{w} + 12 \frac{h}{w} \right]^{-1/2},$$  \hspace{1cm} (3.22)

and the length extensions from the fringing field can then be approximated by [14]

$$\Delta L = 0.412h \frac{\left( \varepsilon_{reff} + 0.3 \left( \frac{w}{h} + 0.264 \right) \right)}{\left( \varepsilon_{reff} - 0.258 \left( \frac{w}{h} + 0.8 \right) \right)}.$$  \hspace{1cm} (3.23)
3. Calculate the resonate length of the microstrip

The resonate length with the length extensions subtracted can then be calculated by [14]

\[ L = \frac{1}{2f_r \sqrt{\mu_r \varepsilon_r \varepsilon_{reff}}} - 2\Delta L. \]  

(3.24)

4. Determine the impedance and probe location

The input impedance for different probe locations was performed using FDTD software. A probe location which will produce a 50 ohm input impedance into the patch can be found iteratively.

5. The resulting field pattern is [14]

\[ E^E_{\phi \text{-plane}} = jk_o W \varepsilon_r e^{-jk_o r} \frac{\varepsilon_r}{\pi r} \text{sinc}\left(\frac{k_o h}{2} \cos(\phi)\right) \text{cos}\left(\frac{k_o L_{eff}}{2} \sin(\phi)\right) \] 

(3.25)

and

\[ E^H_{\phi \text{-plane}} = jk_o W \varepsilon_r e^{-jk_o r} \frac{\varepsilon_r}{\pi r} \text{sin}(\theta) \text{sinc}\left(\frac{k_o h}{2} \text{sin}(\theta)\right) \text{sinc}\left(\frac{k_o L_{eff}}{2} \text{cos}(\theta)\right), \] 

(3.26)

where

\[ \text{sinc} (x) = \frac{\sin(x)}{x}. \]  

(3.27)
3.4.2 Horn Antenna

The pyramidal horn is widely used for making experimental measurements since the gain of the antenna can be made arbitrarily large as a function of its size. One example of pyramidal horn is shown in Fig. 3.8. The gain, $G_o$, of an antenna is directly related to the directivity, $D_o$, of the antenna by the following formula

$$G_o = \varepsilon_o D_o,$$

where $\varepsilon_o$ is the loss factor of the antenna system. The large gain of the pyramidal horn causes a narrow beam pattern for the transmitting and by reciprocity the receiving field pattern. The narrow pattern allows for accurate measurements of transmitting and receiving antenna field patterns in particular at their null locations. In terms of signal processing, the use of the horn antenna measurement can be viewed as a convolution operation. The field pattern of a narrow beam horn antenna can be approximated by a delta function and thus the measured field pattern convolved with a delta function will return the measured field pattern. The approximation of a delta function depends on the directivity of the horn antenna. The gain of the antenna, under lossless conditions, is directly proportional to the physical area of the horn antenna, and is given by [14]

$$G_o = \frac{1}{2} \frac{4\pi}{\lambda^2} (a_i b_i).$$

(3.29)

where $a_i$ and $b_i$ are the outer dimensions of the face of horn. Assuming that the dimensions of the horn $a_i, b_i$ and the dimensions of the rectangular feed $a, b$ satisfy the following condition [14]

$$\left(\sqrt{2\chi} - \frac{b}{\lambda}\right)^2 (2\chi - 1) = \left(\frac{G_o}{2\pi} \sqrt{\frac{3}{2\pi \sqrt{\chi}} - \frac{1}{\lambda}}\right)^2 \left(\frac{G_o^2}{6\pi^3} \frac{1}{\chi} - 1\right),$$

(3.30)
where
\[
\frac{\rho_e}{\lambda} = \chi \tag{3.31}
\]
and
\[
\frac{\rho_b}{\lambda} = \frac{G_o^2}{8\pi^3 \left( \frac{1}{\chi} \right)} \tag{3.32}
\]
the corresponding dimensions of the antenna can then be iteratively solved for by using the following design technique.

1. Pick a desired value of $G_o$, and iteratively solve for $\chi$
2. Determine the corresponding values of $\rho_e$ and $\rho_b$
3. Find the corresponding values of $a_1$ and $b_1$ by using the following equation
\[
a_1 \approx \sqrt{2\lambda \rho_b} = \frac{G_o}{2\pi} \sqrt{\frac{3}{2\pi\chi} \lambda} \tag{3.33}
\]
and
\[
b_1 \approx \sqrt{2\lambda \rho_e} = \sqrt{2\chi \lambda}. \tag{3.34}
\]

Fig 3.8: Picture of a pyramidal horn antenna with dimensions.
3.4.3 Array Design

The array was designed at an operating frequency of 2.424 GHz. The array consisted of a rectangular 3 by 3 array. The size of the array was based solely on manufacturing abilities available and the shape was chosen based on the ease of mathematical formulations. The individual antennas were spaced half a lambda apart in order to avoid spatial aliasing, grating lobes, and to decrease the effects of mutual coupling. With 3 elements centered along the x-axis and 3 elements centered along the y-axis, there are a total of K=3x3 elements. Elements can be counted by rows with a single index \( k = 1 \ldots 9 \), so that \( i_k = 1 + \text{mod}(k-1,3) \), and \( j_k = 1 + \text{int}((k-1)/3) \) [13]. Including mutual coupling effects, the electric field can be described by the following

\[
E(\theta, \phi) \propto E_{\text{element}}(\theta, \phi) \sum_{n=1}^{9} A_n + \sum_{m=1}^{9} A_m S_{nm} e^{j(i_n-2)u + (j_n-2)v}, \tag{3.35}
\]

where \( E_{\text{element}} \) is the dominate polarization of the field pattern [13], \( A \) is the elemental excitation,

\[
u = kd \sin \theta \cos \phi \tag{3.36}
\]

and

\[
v = kd \sin \theta \sin \phi. \tag{3.37}
\]
3.4.4 Mutual Parameters

Assuming the same coupling solution as stated in the active impedance section for a two element array, the field pattern can be written in the following form

\[
E(\phi, S_{11}, S_{12}, S_{21}, S_{22}) \propto E_{\text{element}} \left\{ (A_1 + A_4 S_{11} + A_2 S_{12}) e^{i(\frac{kd}{2})\sin \phi} \\
+ (A_2 + A_2 S_{22} + A_4 S_{21}) e^{i(-\frac{kd}{2})\sin \phi} \right\}.
\]  

(3.38)

Using the objective function [22]

\[
f(S_{11}, S_{12}, S_{21}, S_{22}) = \int_0^{2\pi} \left( E(\phi) - E(\phi, S_{11}, S_{12}, S_{21}, S_{22}) \right)^2 d\phi,
\]

(3.39)

where \( E(\phi) \) is the measured field pattern with known excitation coefficients, one can minimize the squared error using optimization techniques and solve for the unknown scattering parameters \( S_{11}, S_{12}, S_{21}, \) and \( S_{22} \) [35], [36]. A uniform random search was chosen for this work due to its ease of implementation. A genetic optimization or a gradient search may produce more accurate results and be computationally more efficient. The minimization should be performed at multiple scanning angles to ensure the correct solution.

3.4.5 Corrective Network and Scan Blindness

Assume that the electric field pattern can be described by the following formula

\[
E_{\phi}^{\text{coupled}} \propto E_{\text{element}} \left\{ (A_1 + A_4 S_{11} + A_2 S_{12}) e^{i(\frac{kd}{2})\sin \phi} \\
+ (A_2 + A_2 S_{22} + A_4 S_{21}) e^{i((-\frac{kd}{2})\sin \phi)} \right\},
\]

(3.40)

and that the desired field pattern can be described with the following coefficients
The arrangement of the array is as shown in Fig 2.5. The coefficients can then be presented in matrix form

\[ \begin{bmatrix} C_1 \\ C_2 \end{bmatrix} = \begin{bmatrix} 1 + S_{11} & S_{12} \\ S_{21} & 1 + S_{22} \end{bmatrix} \begin{bmatrix} A_1 \\ A_2 \end{bmatrix} \Rightarrow \begin{bmatrix} A_1 \\ A_2 \end{bmatrix} = \begin{bmatrix} 1 + S_{11} & S_{12} \\ S_{21} & 1 + S_{22} \end{bmatrix}^{-1} \begin{bmatrix} C_1 \\ C_2 \end{bmatrix} \]  

(3.42)

where \( A_1 \) and \( A_2 \) are the complex input excitations, and \( C_1 \) and \( C_2 \) are the desired complex output excitations. The inversion can be accomplished by using Cramer’s rule. The input excitations can then be solved for and implemented using phase shifters and adjustable amplifiers as shown in Fig. 3.9. \( A_1 \) and \( A_2 \) are given by

\[ A_1 = \frac{(1 + S_{22})C_1 - S_{12}C_2}{(1 + S_{11})(1 + S_{22}) - S_{12}S_{21}}, \quad A_2 = \frac{(1 + S_{11})C_2 - S_{21}C_1}{(1 + S_{11})(1 + S_{22}) - S_{21}S_{22}}. \]  

(3.43)

The scattering parameters can be obtained either through direct measurement [31, pp.61-80] through the method of open and short circuits [37, pp.124],

\[ Z_{oc} = Z_{11} = Z_{22}, \quad Z_{12} = \sqrt{Z_{oc} (Z_{oc} - Z_{sc})}, \]  

(3.44)

or through far field measurements [13]. Scan blindness, resulting in a lack of radiation or reception in a particular direction, can occur during the following conditions

\[ \begin{bmatrix} 1 + S_{11} & S_{21} \\ S_{12} & 1 + S_{22} \end{bmatrix} \begin{bmatrix} A_1 \\ A_2 \end{bmatrix} = \begin{bmatrix} 0 \\ 0 \end{bmatrix} \Rightarrow \frac{A_1}{A_2} = \frac{-S_{21}}{1 + S_{11}} \quad \text{and} \quad \frac{A_2}{A_1} = \frac{-S_{12}}{1 + S_{22}} \]  

(3.45)

or

\[ (1 + S_{22})(1 + S_{11}) = S_{12}S_{21}. \]  

(3.46)
This means that the excitation is completely independent of the scanning angle. However, this condition has not been found to be true experimentally [21]. The use of a higher order model should lead to a solution which can better predict scan angle blindness.

3.4.6 Automatic Tuner

Fig 3.9: A diagram indicating possible elements used in a mutual coupling corrective network

Fig 3.10: A diagram indicating a network implementation of an automatic tuning alignment system for mismatches between the power amplifier and the antenna.
Due to the active impedance model, any reflections caused by mismatch in impedance due to manufacturing variations will disrupt the field pattern, reduce the overall gain of the system and disrupt the VCO (RF power source) as depicted in Fig. 3.10. Therefore, a structure was proposed that automatically aligns the matching network to minimize the reflected energy that is caused by a mismatch in the impedance of the antenna to the system impedance that appears at the antenna terminals. The automatic tuner system proposed can be described by four parts: a matching network, a coupler, power detector, and an algorithm. The matching network could be performed using a standard $Q^2 + 1$ match that is matched for frequency and varied by varactors. A directional coupler is used to measure the reflected energy caused by the mismatch at the antenna terminals. The coupler that is usually built using coupled microstrip lines can alternatively be done using discrete components as shown in Fig. 3.11. Trades offs exist between increased component cost, size, directivity, and decreased bandwidth.

The design procedure can be broken into two steps [31]

1. Input Impedance and Coupling Coefficients
   \[ Y_0 = \sqrt{Y_{0e}Y_{0e}} \]  
   \[ c = \frac{Y_{0o} - Y_{0e}}{Y_{0o} + Y_{0e}}. \]  

2. Component Values
   \[ L_{21} = \frac{2}{\alpha(Y_{0o} + Y_{0e})}. \]
The power detector used to generate an objective function measures the amount of reflected energy. The gradient algorithm then adaptively adjusts the matching network to minimize the objective function. A gradient algorithm was chosen based on the author’s previous experience using optimization algorithms to design microwave networks. The algorithm leads to a fairly robust solution, due to its continuous nature.

![Diagram of a discrete version of a directional coupler](image)

Fig 3.11: A diagram that shows a structure of a discrete version of a directional coupler [31].

\[
C_{11} = \frac{1}{\omega^2 L_{21}} - C_{13} = \frac{Y_{0e}}{\omega}, \quad (3.50)
\]

and

\[
C_{13} = \frac{Y_{0e} - Y_{0o}}{2\omega}, \quad (3.51)
\]

with

\[
C_{11} = C_{22} = C_{33} = C_{44}. \quad (3.52)
\]
3.4.7 Modulation Format

It will be shown that the phased array pattern is independent of a given modulation scheme assuming no carrier frequency change [27]. For example, a QPSK modulation scheme can be described in terms of the following excitation per symbol

\[
\bar{A}_{\text{baseband}}^p = A_p \sin \left[ \omega t + (i - 1) \frac{\pi}{2} + \phi_o^p \right]
\]

(3.53)

for \( i = 1, 2, 3, 4 \) and the excitation coefficients at the antenna terminals can be represented as

\[
\bar{A}_{\text{RF}}^p = A_{\text{mod}} A_p e^{j[(i - 1) \frac{\pi}{2} + \phi_o^p]}. \tag{3.54}
\]

---

**QPSK Constellation**

**QAM Constellation**

![QPSK and QAM Constellations](image)

Fig 3.12: A diagram showing a QPSK modulation (Left) and QAM modulation in terms of its real and imaginary components (Right).
The field pattern is then related to the element excitation by Eq. 3.38. Inserting the excitation into an active field pattern, and after factoring the common term, \( A_{\text{mod}} e^{j[(i-1)\pi/2]} \), from the excitations, the field pattern can be written as

\[
E_{\phi} \propto E_{\text{element}} \left\{ \left( A_1 + A_1 S_{11} + A_2 S_{12} \right) e^{j(kd/2)\sin \phi} + \left( A_2 + A_2 S_{22} + A_4 S_{21} \right) e^{j(-kd/2)\sin \phi} \right\} A_{\text{mod}} e^{j[(i-1)\pi/2]},
\]

which is independent of the modulation angle. The \( S_{ij} \) are scattering parameters from port \( i \) to port \( j \). This can be generalized to any modulation scheme. The architecture presented is more suitable for QAM and QPSK modulations shown in Fig. 3.12. The bit rates of the system will be limited by the convergence of the phased locked loop, which is dependent on the loop’s gain and bandwidth. There exists a tradeoff between the convergence of the loop and the amount of phase noise that is introduced by the phased locked loop due to the bandwidth of the loop. A phase aid can be used to mitigate the tradeoff between convergence and noise. Bruce Fu (a colleague working on the system) will be continuing to work with phase aiding in his thesis.
3.4.8 Polarization Adjustment

Typically, MIMO configurations include the use of orthogonal polarization antennas to increase the diversity gain of the system and the use of switches [38-40]. This type of scheme does increase the probability of receiving a transmitted signal; however, a maximum 3 dB loss can occur over all available polarization of the received signal using such a diversity technique. Additional antennas can be added at arbitrary polarizations to overcome the loss, but hardware cost increases as a function of the number of additional antennas. Three different types of polarizations (linear, circular, and elliptical) can be excited by a microstrip antenna shown in Fig. 3.13. Any wave can be broken down into right handed waves and left handed waves [41]. This can be demonstrated using a linear polarized field [41]

Fig 3.13: A diagram showing a microstrip patch antenna configuration in terms of a measurement axis.
Assume that the wave can be broken down into left handed and right handed circular polarized waves represented as

$$\mathbf{E}(x) = \mathbf{A}_l(x) + \mathbf{A}_r(x).$$  \hspace{1cm} \text{(3.57)}

A right handed wave takes the following form:

$$\mathbf{E}_r = \frac{E_0}{2} \left( \mathbf{a}_y - j \mathbf{a}_z \right) e^{-j k x}.$$  \hspace{1cm} \text{(3.58)}

A left handed wave takes the following form:

$$\mathbf{E}_l = \frac{E_0}{2} \left( \mathbf{a}_y + j \mathbf{a}_z \right) e^{-j k x}.$$  \hspace{1cm} \text{(3.59)}

Substituting, these into Eq. 3.38 gives

$$\mathbf{E}(x) = \mathbf{E}_r(x) + \mathbf{E}_l(x). = \mathbf{A}_l e^{-j k x}.$$  \hspace{1cm} \text{(3.60)}

These procedures can be generalized to any mixture of polarizations by properly weighting a linear combination of right and left handed polarized waves. The time average power received can then be represented as

$$P_{\text{received}} \propto |\mathbf{a}_{rx} \cdot \mathbf{a}_{rx}|^2 = \cos^2(\theta_{TR}),$$  \hspace{1cm} \text{(3.61)}

where $\theta_{TR}$ is the angle in between the receiving and transmit vector as shown in Fig. 3.14. The orthonormality of left handed and right handed polarized waves allows one to form a basis that an auto-alignment system can be built [42]. The maximum misalignment occurs at
90 degrees. This is referred to as an orthogonal polarization and results in zero power reception. A probability distribution can be assumed as a random uniform variable with a distribution as shown as

\[
f(\theta_{TR}) = \begin{cases} 
\frac{2}{\pi} & 0 < \theta_{TR} < \frac{\pi}{2}.
\end{cases}
\tag{3.62}
\]

The SNR is a function of the polarization alignment

\[
SNR = \frac{\langle x(t)^2 \rangle}{\langle n(t)^2 \rangle} = \frac{\sigma_{message}^2}{\sigma_{noise}^2} E[\cos^2(\theta_{TR}(t))],
\tag{3.63}
\]

assuming a zero mean signal, and independence between the polarization of the signal and the signals content. The average degradation factor of the signal-to-noise ratio, SNR, due to misalignment can then be calculated by taking the expectation value of the SNR

\[
\bar{\gamma} = SNR_{max} \int_{0}^{\pi} \frac{2\theta_{TR}}{\pi} \cos^2(\theta_{TR}) \, d\theta_{TR} = \frac{2}{\pi} \left[ \frac{\pi^2}{16} - \frac{1}{4} \right] SNR_{max},
\tag{3.64}
\]

\[\hat{a}_{Tx} - \theta_{TR} - \hat{a}_{Rx}\]

Fig 3.14: Polarization unit vectors of the transmitter and receiver in terms of the polarization loss factor. [14]
where

\[
SNR_{\text{max}} = \frac{\sigma_{\text{message}}^2}{\sigma_{\text{noise}}^2}.
\]  

(3.65)

Similarly, channel capacity can be defined as

\[
C = \log_2[1 + SNR],
\]

(3.66)
assuming a constant signal-to-noise ratio. Treating the SNR as a random variable, the average degradation of the channel capacity can be represented as

\[
\bar{C} = \int_{0}^{\pi/2} \frac{2\theta_{TR}}{\pi} \log_2[1 + SNR_{\text{max}} \cos^2(\theta_{TR})] d\theta_{TR}.
\]

(3.67)

Multi-path fading occurs due to reflections from objects and consists of a dominate line-of-sight (LOS) component and a diffuse non-line-of-sight (NLOS) component [43]. The NLOS component’s polarization changes from the dominate LOS component due to reflections from obstacles [44]. Therefore, the received signal can be represented as a summation of varying polarized NLOS components and a dominate LOS component as was shown in Fig. 1.1. An automatic alignment scheme can be implemented by adding extra bits in the preamble of the transmitted data. The bits either represent that the signal strength has increased or decreased. Searching of the alignment space can be done using the binary chop algorithm with a maximum computational time varying as log(N) or through a linear search algorithm with a maximum computation time depending on N where N is the number of divisions in between 0 and 90 degrees. The advantage of the binary chop algorithm, shown in Fig. 3.15, is that half the search space is discarded with each estimate of the signal. A
disadvantage of the algorithm is the fact that multi-path fading or white noise can cause the
algorithm to improperly discard the interval containing the correct polarization. With the
linear search, shown in Fig. 3.16, computational time is increased but performance under
multi-path fading and noisy environments should be more robust. In order to achieve zero
loss due to polarization misalignment, the antenna’s polarization vectors must be exactly
aligned or in parallel. This can be accomplished by weighting the signals into the feeds of the
antenna.

Fig 3.15: A flow diagram of a binary search algorithm.
The field can then be represented by

$$E_{\text{total}} \propto A_{\text{feed}1}(\hat{\alpha}_y + j\hat{\alpha}_z) + A_{\text{feed}2}(\hat{\alpha}_y - j\hat{\alpha}_z),$$

(3.68)

which will produce a right handed,

$$E_{\text{righthanded}} \propto A_{\text{feed}1}(\hat{\alpha}_y + j\hat{\alpha}_z),$$

(3.69)

and a left handed,

$$E_{\text{lefthanded}} \propto A_{\text{feed}2}(\hat{\alpha}_y - j\hat{\alpha}_z),$$

(3.70)

circularly polarized wave weighted by the excitation feed, assuming the architecture shown on the right in Fig. 3.17 and

$$E_y = E_z.$$  
(3.71)
This assumption will occur under symmetrical conditions. The alignment can be accomplished by adjusting the weights and phases of the signals shown on the left in Fig. 3.17 and given by

\[ E_{\text{total}} \propto A_{\text{feed}1} \hat{a}_z + A_{\text{feed}2} \hat{a}_z. \]  \hspace{1cm} (3.72)

where \( A_{\text{feed}1} \) and \( A_{\text{feed}2} \) are the complex feed excitations.
3.5 Summary

All previous results were judged on implementation, cost effectiveness, and manufacturability following current and future industrial trends. The author acknowledges that future trends may change based on innovation but remains firm on his criteria. A novel phase shifter using a shift register and phase locked loop was shown which has advantages over traditional techniques in cost, repeatability, linearity, and resolution capabilities. The disadvantage of the structure is that the resolution is limited by a master clock and sensitivity of the digital division used in most PLLs. An offset PLL was then introduced to remedy this sensitivity issue. The mutual coupling theory was derived from Pozar’s work [13] and show that the baseband pre-distortion techniques can theoretically correct for mutual coupling. The pre-distortion technique will only enhance the results of hardware mutual reduction methods. An inconsistency was also shown with scan blindness and the current theory of mutual coupling. A benefit of a baseband phase control methodology is that it is guaranteed not to disturb the fields of the antenna that are based on the boundary conditions surrounding the antenna. With the addition of more hardware, it may be possible to compute the pre-distortion phases dynamically to mitigate the effect of external reflections. A new measurement technique for measuring coupling parameters based on an optimization routine on the far field pattern was introduced which will reduce testing time in large arrays. An adjustable amplifier was also shown with an advantage over previous structures based on the fact that zero phase distortion is possible. Phase aid using charge injection reduces complications of stability of the phased locked loop, charge injection caused by switches, and reduces the filters dependency on the lock detect. The automatic tuner derived in this
dissertation and found in previous work is very simple; the works were however developed independently [28]. The primary difference includes the type of coupler, tuning element and algorithm used. In terms of the coupler used, there will be a tradeoff between cost, directivity, and bandwidth. A more detailed analysis of the difference of the tuning element such as linearity, convergence and tuning range will be left for future work. The analysis of the modulation types shows that the platform is capable of supporting a novel result. The polarization method using excitation of different modes within the patch is more elegant compared to other methods based on cost factors for adding additional antennas in a diversity scheme and it will not induce harmonics such as a varactor being used directly on the antenna would. Overall, new theoretical ideas were presented which will only enhance the field of the antenna design through cost savings, better accuracy, and new methodologies.
CHAPTER 4  EXPERIMENTAL SETUP

4.1 Introduction

An experimental setup was implemented to expedite and increase the accuracy of the previous antenna setup at Iowa State University. An automated network using a coherent network analyzer and a non-coherent spectrum analyzer in combination with an optical detector is presented. This will increase the accuracy of the measurements and expedite the measurement results. Filters and a measurement setup are presented to allow for future continuation and repeatability of the work presented in this dissertation.

4.2 Setup

In the initial experimental setup, a two element dipole and a microstrip antenna arrays were used to work out initial flaws and to verify theoretical results that were previously published.

Fig 4.1: An automatic test setup for non-coherent measurement of an array
This antenna setup is shown in Fig. 4.1 and Fig. 4.2, and was later expanded to the 3x3 phased array. Fig 4.1 allows for a non-coherent detection whereas Fig 4.2 allows for a coherent measurement system. The relevancy of coherency is determined by whether one needs carrier phase information from the system. As proposed by Pozar [13], the mutual coupling parameters can be gathered solely from the far field measurements which require phase information or coherency of the wave front at various locations in space. For the noncoherent system, the spectrum analyzer is connected to a computer which synchronizes the instruments to an angular rotary device. The spectrum analyzer is configured for narrow band measurements that are averaged to reduce measurement variation by the square root of the average factor. The reduction in variation allows for low side lobe measurements to be performed. The exact phase differences, $\phi_2 - \phi_1$, and amplitudes, $C_1$ and $C_2$, between the input signals were measured using an oscilloscope and these signals can be described by the equations below:

$$V_1(t) = C_1 \sin(\omega_f t + \phi_1),$$

(4.1)

and

$$V_2(t) = C_2 \sin(\omega_f t + \phi_2).$$

(4.2)

Fig 4.2: An automatic test setup for a coherent test setup.
Scattering parameter measurements of the amplifier, filter and interconnecting cables were combined to obtain

\[ V_{p, \text{antenna}}(t) = C_{p, \text{antenna}} \sin(\omega_{\text{rf}} t + \phi_{p, \text{antenna}}) \]  

(4.3)

where

\[ C_{p, \text{antenna}} = G_{p, \text{filter}} G_{p, \text{amp}} G_{p, \text{cable}} A_p \]  

(4.4)

and

\[ \phi_{p, \text{antenna}} = \phi_{p, \text{cable}} + \phi_{p, \text{filter}} + \phi_{p, \text{amp}} + \phi_0^p \]  

(4.5)

for \( p = 1, 2 \) using the properties of a linear system. The scattering parameters of the array are directly measured and combined with the active field pattern to predict field pattern measurements. The phase shifts shown were originally performed using transmission lines, shown in Fig. 4.3, for time delays. These were later replaced by phased locked loops as phase shifters.

Fig 4.3: A phase shifter using microstrip delay lines
The delay lines were formulated to provide an incremental phase shift between 0 and 180 degrees. A resistive splitter was designed and is shown in Fig. 4.4. The resistive architecture was chosen over the lower insertion loss architecture primarily due to the compactness of the design. The purpose of the splitter is to provide uniform phase and amplitude excitations to both of the elements. Harmonics at multiples of the VCO’s output frequency will be generated in the phased locked loop and power amplifiers due to nonlinear effects of the transistors, and saturation effects from the oscillation loop gain of the VCO being greater than one. The harmonics may in turn distort the field pattern of the array due to the fact that antennas, depending on their type, are typically

Fig 4.4: A resistive splitter for equal amplitude and phase excitation of the array
resonant at multiples of the fundamental design resonance. These harmonics are of particular concern compared to other spurious emissions due to the fact that patch antennas will resonate at multiples of the design frequency. In order to mitigate the harmonic effects, a low pass filter was placed in front of the phased locked loop with the pass band containing the fundamental and the stop band containing the harmonics. A low pass architecture shown in Fig. 4.5 was chosen over band pass architecture due to the fact that transmission line based band pass filters have a tendency to resonate at multiples of the design frequencies. The transmission line designs were chosen based on the ease of manufacturing, debugging purposes, and good agreement between experimental and simulation results. The measured results are shown in Fig. 4.6.

Fig 4.5: A stub low pass filter used for harmonic removal
Fig 4.6: Stub low pass filter’s magnitude response

The insertion loss of the filter is negligible for the purpose of this study and has 40 dB of attenuation at the 2\textsuperscript{nd} harmonic frequency. A band pass filter shown in Fig. 4.7 was placed

Fig 4.7: A band pass filter
behind the spectrum analyzer in order to limit thermal noise content and response to interfering signals that were received by the horn antenna. The measured results are shown in Fig. 4.8. Note that the filter resonates at the fundamental design frequency and at a multiple of the primary resonance. At higher frequencies, the filter has a high pass filtering response. This can be explained by increased coupling and surface wave effects.

Fig 4.8: The measured response of the bandpass filter

Fig 4.9: Coupled line filters
Additional couple line filters, as shown in Fig. 4.9, were fabricated and tested. However, the author was not satisfied with their experimental results and they were thus discarded from the design. Power amplifiers chosen for the design were Mini-Circuits® lee_39. The amplifiers were chosen based on their relatively high 1 dB compression point of 18 dBm and their broad band gain up to 10 GHz. They can be used in other applications such as for an automated tester for an anechoic chamber. The lab setup uses a receiving horn antenna with about 28 dB of gain. This allows for a high degree of measurement accuracy for the transmitting antenna. The rotary motor for the transmitting antenna was then synchronized with the spectrum and network analyzer to make automated power measurements. The distance between the transmitting and receiving antennas was such that far field pattern measurements were taken. The lab setup is shown in Fig. 4.10. Optical cables, such as shown

Fig 4.10: Automatic test setup consisting of horn antennae in the (Upper Left), array in the (Upper Right), synchronous step rotary (Lower Left), and spectrum analyzer (Lower Right).
in Fig. 4.11, were used with a bar code strip to synchronize the rotary motor and the measurements taken with the network analyzer. The critical angle of propagation for the optical decoder to determine the positioning of the cables as shown in Fig. 4.11 is [46]

\[
\alpha_c = \sin^{-1} \sqrt{1 - \left( \frac{n_2}{n_1} \right)^2},
\]  

(4.6)

where \( n_1 \) and \( n_2 \) are the indices of refraction of the core and cladding respectively and \( n_1 \geq n_2 \). This can be related to the acceptance angle, \( \phi_a \), by [46]

\[
n_a \sin(\phi_a) = n_1 \sin(\alpha_c),
\]  

(4.7)

![Figure 4.11: Optically encoded turn where the black marker indicates zero and white marker indicates a one [46].](image-url)
where \( n_a \) is the index of refraction of the air. The cables were experimentally positioned so that the counted number of serial bars were repeatable at various speeds of automation. A measurement was taken by the spectrum analyzer every time a white marker was detected. The measurements could then be correlated with the relative position of the markers.

### 4.3 Summary

The test setup was judged on the ability to repeat the experimental results, accuracy of the measured data, and the reduction of testing time (relative to the previous laboratory setup). The resolution of the automatic test setup using an optical encoder allowed for measurements of less than a degree to be taken, which is better than the spatial resolution of the horn antenna. To reduce the noise of the system, the antenna was rotated and the measurement results were averaged. This led to results that were measurable up to the accuracy of the spectrum analyzer, which is approximately 1/6 dB. In summary, the test setup allowed for repeatable and accurate measurements, only limited by the accuracy of the test equipment itself. Overall, the measurement system was a success and allowed for rapid accurate measurements.
CHAPTER 5  EXPERIMENTAL AND SIMULATION RESULTS

5.1 Introduction

Experimental results are presented for the phase shifters using a scaled version of the hardware phase shifter. The accuracy of the active impedance model for a 2x1 microstrip array and a dipole array is compared against experimental results, the standard model with and without mutual coupling effects. Steering was demonstrated in the 2x1 dipole array and a 3x3 microstrip array using the hardware phase shifter units. Simulation results of the autotuner and the polarization alignment scheme are presented that allow for a more optimal system.

5.2 Phase Shifter

5.2.1 Time Delay

The results of the waveform shifted at baseband versus the RF phase shift are shown in Fig. 5.1. The correlation between baseband and RF phase shift was stated in Eq. 3.3.

Fig 5.1: A baseband phase shift at 3.125 MHz (Left) and an RF phase shift at approximately 2.7 GHz (Right).
Original measurements were made for the phase shift without scaling taken into account and thus the waveform at RF and baseband differed by the divide ratio multiplication. Fig. 5.2 shows the results of the percent error with the multiplication factor taken into account. The accuracy limited to phase noise reduction, which was reduced using ensemble averaging techniques. However, there was an anomalous point which the author assumes is a cycle slip in the d-flip flop, due to unequal time delays in the multiplexer paths.

After the correction was made to remove the multiplication factor from Eq. 3.3, the results shown in Fig. 5.3 appear to be very exact, except at 2 major discrepancy points that again the author assumes that there were cycle slips. This would account for the symmetry of the points in the graph. In conclusion the results appear to match the theoretical analysis accurately, with a minimal contamination due to phase noise error and cycle slips in the logic.
5.3 Phased Locked Loops

5.3.1 Offset Mixers

Fig 5.3: Theoretical versus experimental phase shift for the phase shifter with an RF at 2.53125 GHz, a baseband at 3.125 MHz and with a phase detector at 312.5 KHz.

Fig 5.4: A fabricated phased locked loop using offset mixers for frequency conversion of the RF down to the reference frequency.
A phased locked loop with offset mixers was successfully fabricated and is shown in Fig. 5.4. The output lock frequency is shown in Fig. 5.5. The locking range of the device was approximately 2 MHz. A coherent source needs to be utilized in order to verify that there is no phase scaling from the input reference to the output reference. The device was not integrated into the phased array due to the number of discrete components needed and noise issues. Implications of this structure should be investigated in the future in terms of phase tracking.

5.4 Antenna and Arrays

5.4.1 Antenna

The simulated directivity and the gain of a single patch element used are listed in Table 5.1. This is typical of a microstrip patch antenna on a 30 mil board with a dielectric constant of 3.38.
Better improvement in gain and directivity can be improved by better impedance matching and by varying the board thickness respectively. Fig. 5.6 shows a HFSS simulated pattern of the patch.

Table 5.1: HFSS simulated gain and directivity of microstrip patch at 2.425 GHz [45].

<table>
<thead>
<tr>
<th>Directivity Db</th>
<th>Gain Db</th>
</tr>
</thead>
<tbody>
<tr>
<td>5.56</td>
<td>4.26</td>
</tr>
</tbody>
</table>

Fig 5.6: The electric field strength emanated by a microstrip patch antenna at 2.425 GHz [45].
5.4.2 Array Design

A 3x3 phased array was fabricated given the previous architecture and a picture of the array is shown in Fig. 5.7 as it was installed in the anechoic chamber. In order to calibrate the system output phases and amplitudes of each element were measured by removing the antenna array, which allowed direct measurement of the outputs of the individual VCO’s.

Fig 5.7: A functional 3x3 phased array.
It will likely be necessary to integrate the phased array onto a single PCB board to minimize coupling effects. However, sufficient control of the experimental phased array was accomplished to demonstrate steering of the array. The coupling between wires in the external wiring, shown in Fig. 5.8, affects the phases on each antenna. Any slight capacitive delay or cross coupling effect at baseband gets multiplied by the scaling factor. If sufficient phase shifting resolution were obtained by the clock, then a possible solution would be to calibrate the system and use pre-distortion to account for variations in wiring. The results of a two by one dipole array correspond with Eq. 2.16. as demonstrated by measurements shown in Fig. 5.9. The discrepancy in shape can be partially explained by finite ground plane effect, or edge effect and other distortions such as wobble of the measurement system turn wheel. The measured results should in theory be able to be extrapolated up for larger arrays, where mutual coupling will become more prominent.
The results indicate that a large array can be completely controlled by a phased locked loop. The measured phase shift for the 3x3 antenna is shown in Fig. 5.10. This demonstrates the ability to perform phase steering with the current circuitry. The circuitry would need to be improved before a correlation between the experimental and active impedance model and pre-distortion technique could be better verified.

Fig 5.9: Theoretical verses experimental field patterns of a 2x1 dipole array operating with RF at 2.425 GHz and baseband at 3.125 MHz for two different phase shift patterns.

The 3x3 phased array measurements with an RF at 2.53125 GHz, a baseband at 3.125 MHz and a phase detector at 312.5 KHz. Two different phase shifts were demonstrated.

Fig 5.10: The 3x3 phased array measurements with an RF at 2.53125 GHz, a baseband at 3.125 MHz and a phase detector at 312.5 KHz. Two different phase shifts were demonstrated.
Fig 5.11: A diagram showing the tuning capabilities of the automatic tuner for frequencies outside the resonant frequency for two microstrip patches that are coupled.

5.4.3 Automatic Tuner

Fig. 5.11 shows the reflection coefficient, $S_{11}$ and $S_{22}$, for two mutually coupled antennas depicted in Fig 3.10 that are tuned to a resonant frequency through the use of the steepest decent algorithm. The antennas were originally designed for resonant frequency of 2.425 GHz. The capacitance in the $Q^2 + 1$ match were varied to minimize the error (reflected energy) that is caused by a mismatch in the impedance of the antenna. Using a typical varactor tuning range as a constraint and 15 dB of directivity on each coupler, the auto-tuner was able to adaptively tune over a 200 MHz bandwidth. This would well meet the demands of typical misalignment due to manufacturing errors and also meet most frequency hopping applications requirements. When using varactors, placement of the varactors should be made with extreme caution. Careful placement will minimize the harmonics generated by varactors.
due to high output power coming from the antenna. Other more suitable techniques may exist. Other practical issues that may exist include the effects of interfering received signals, fixed point operation of a DSP or FPGA, thermal noise, and energy reflected back towards the antenna by external objects. A coupler was successfully fabricated and is shown on the left in Fig. 5.12. The measured results of the coupler are shown on the right in Fig. 5.12. These results meet the bandwidth requirement of 200 MHz with a directivity of 14 dB.

### 5.4.4 Polarization Adjustment

Simulations results, as shown in upper left of Fig. 5.13, show that the linear search used takes longer to converge as compared to the binary chop algorithm used under a worst case scenario when the polarization of the transmitting and receiving antennas are orthogonal. The faster convergence of the binary chop algorithm is due to the fact that half the search space is discarded with each sample point. The lower of Fig. 5.13 shows the performance of both algorithms under different SNR environments. The linear search algorithm performs worse under extreme noisy conditions as compared to the binary chop.
Fig 5.13: The figures show the convergence results of binary chop and a linear search algorithm (Upper Left). The figure also contains the convergence of algorithms using a multipath, 20 uniform scatterer environment (Upper Right). The results for different SNR environment for both algorithms (Lower).
algorithm. This can be attributed to the fact that under extreme noisy environments, the binary chop algorithm has an equal probability of discarding the correct interval or choosing the incorrect interval. However, the linear search algorithm performs better at a higher SNR due to the fact that the binary chop algorithm has the possibility of discarding the wrong interval while the linear search algorithm will result in an approximate region of the correct polarization. The upper right diagram of Fig. 5.13 shows similar trends as does the lower of Fig. 5.13 that was simulated under a multi-path environment. The multi-path environment was approximated by twenty scatterers using a random uniform variable for the polarization component.

5.4.5 Mutual Coupling

Fig. 5.14 shows a difference between the experimental results and the standard dipole array model. Improvements are found when the active impedance, Eq. 2.16, is taken into account. The most noticeable differences in the field pattern can be seen at lower levels of the field pattern. This can be attributed to the domination of the coupling parameter terms.
\[ A_1 S_{11} + A_2 S_{12} \quad \text{and} \quad A_2 S_{22} + A_1 S_{21}, \]  

where \( A_1 \) and \( A_2 \) are the complex excitation coefficients of the array. The main beam seems to be only mildly affected by the coupling parameters. From the experimental data of a microstrip array shown in Fig. 5.15, there is a slight improvement between the experimental and active impedance models. This can be attributed to smaller coupling parameters. The active impedance model may also be viewed as a result of a first order approximation of a Taylor expansion

\[ E \propto f_1(A_1, A_2)e^{j(kd/2)\sin \phi} + f_2(A_1, A_2)e^{-j(kd/2)\sin \phi}, \]  

with respect to the excitation coefficients. Since there are nonlinear effects in the system, the use of a higher order model may produce a more accurate model of the system. The coefficients can be found by using an optimization routine to minimize the error.

![Element Pattern](image1.png) ![Measured Array](image2.png) ![Array Comparison](image3.png)

Fig 5.15: Experimental (Exp) versus theoretical results using the standard model (Std) and the active impedance model (Act) for a 2x1 microstrip array at 2.425 GHz.
5.4.6 Mutual Coupling Optimization

The results of an optimization, Eq. 3.39, can be seen in Fig. 5.16. There is an improvement over the results in Fig. 5.15 in terms of the low level field patterns. Minimization should be performed at multiple scanning angles in order to verify a correct solution.

5.5 Summary

The experimental results were judged on its correlation to the theory described in chapter 3, as well as whether deviations from the theory could be explained. New ideas were highlighted for theoretical development. Experimental results for the phase shifter followed very close to the theoretical development, both when scaling was present and when it was not present. Discrepancies that were presence were attributed to hardware implementation issues. Using the phased locked loop as phase shifters, the pattern was steered. The automatic tuner simulation
shows that the antenna is capable of being tuned at different frequencies. Future work needs to be done in experimentally proving the automatic tuner result. Using Matlab and its mathematical libraries, the idea of using a dual fed microstrip antenna to automatically align the polarization of receiving and transmitting signals was demonstrated both in the presence of white noise and in a multipath environment. The experimental trends for the field pattern of the array follow the theoretical model to a great deal of accuracy for the dipole array, but showed same deviation from theory for the microstrip array. The inaccuracies of a microstrip experiment can be explained by the existence of higher order modes present in the patch, or due to the presence of a finite ground plane. Mutual coupling parameters were experimentally shown to be obtained through optimization routines of the far field measurement. The final array design demonstrates that the beam is capable of being steered with phased locked loops. Further work needs to be performed in order to perfect the circuitry and the system.
CHAPTER 6  FUTURE WORK, SUMMARY, AND DISCUSSION

6.1 Introduction

The project demonstrates antenna beams can be digitally steered using a FPGA and a phased locked loop to control the phased array. New experimental techniques were developed to allow more precise characterization of the system, which allows for more accurate prediction of the beam. New items for investigation, such as scan blindness, scaling of the input phase, phase noise, impedance variations, and mutual coupling effects in external circuitry, have been shown to cause a deviation from the current theory. The next step for this project would be to develop an integrated circuit using a larger phased array in order to be able to control both the main beam and null locations. Additional techniques were theoretically developed to increase power gain and to resolve manufacturability concerns. New architectures were presented to resolve the scaling factor of the input phase, allowing for a fast hopping phase system. Future work needs to be performed in these areas before a successful implementation of the system into a practical system can be done.

6.2 Overview of Successes

Through experimental data, the system using shift registers and phased locked loops was shown to be able to control the phase and thus the beam of the main antenna system. The theory of an active impedance model was experimentally shown to be verified and a development of a mutual coupling corrective network was developed. Methods for auto-
tuning, amplitude adjustment without phase distortion and auto-polarization alignment were discussed and/or simulated.

6.3 Future Work

Future work should include experimentally building and verifying the method for auto-tuning, amplitude adjustment without phase distortion, and the auto-polarization system. Future work should also theoretically further the active impedance model and scan blindness effects. It is suggested that one build a system on a single PCB board and use computational techniques to further enhance the theoretical versus experimental results.

6.4 Conclusion

Overall, the project had a lot of setbacks and successes. The author personally believes technology needs to mature and become less expensive before integrating into commercial technology and thus mitigating the issues that have cropped up in the external circuitry. At best, the project has laid a solid foundation for both theoretical and experimental analysis.
BIBLIOGRAPHY


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